REMARKS

This Amendment is submitted in response to the Office Action, dated June 21, 2005. Claims 1, 7 and 14 are proposed for amendment herein. Claims 4 and 15 are cancelled without prejudice. Claims 1-3, 5-14 and 16-17 are presently pending in the above-identified application.

Rejection of Claims under 35 USC § 112

The Office Action rejected claims 1-6 and 11-13 under 35 USC § 112, first paragraph, as failing to comply with the enablement requirement. In particular, the Office Action stated that "...with respect to the means for electrically connecting said transmission element to said at least one ground plane, it would have been unclear how one skilled in the art would be enabled to use the invention since the apparent electrical connection of the transmission element to the ground plane would seemingly short circuit (to ground) any signals propagating along the transmission element, thereby negating and rendering inoperative the "transmission line"."

Applicant respectfully disagrees. In particular, as described in Applicant's Specification, as originally filed, the principles of the invention relate to high frequency electronic circuit boards. In particular, such high frequency boards carry signals represented by alternating currents with frequencies above 70GHz (see, e.g., Applicant's Specification, page 3, line 23 through page 4, line 10; and page 1, lines 19-24). As will be appreciated by those skilled in the art, due to their small wavelength, such signals fall into the microwave domain, therefore, standard circuit theory cannot be applied directly. That is, in these microwave applications, the device dimensions are on the order of the microwave wavelength and the phase of a voltage or current changes significantly over the physical length of the device. Importantly, in the special case that the transmission line length equals an uneven multiple of a quarter wavelength, output short circuited (to ground) lines exhibit open circuit behavior at their input (see, e.g., David M. Pozar, Microwave Engineering, Addison Wesley 1990, ISBN 0-201-50418-9, pp. 1-2, 21, and 79-80 copy enclosed herein for reference).

In recognition of these microwave wavelength characteristics, the Applicant herein has realized that the supporting elements of the claimed invention may leverage this phenomenon, in particular, to reduce (or substantially eliminate) dielectric signal loss. That is, in accordance with the amended claims herein, the claimed supporting elements provide two required functions: (1) suspending the claimed transmission element a desired distance away from the ground plane; and (2) electrically connecting such transmission element to the ground plane such that the transmission line is not in electrical contact with a substrate other than through the support elements. At the higher frequencies of operation (i.e., at or above 70 GHz) the claimed supporting elements of the present invention are virtually non-existent from a signal perspective and therefore do not have a deleterious affect on performance of the transmission line/element. That is, the transmission line/element of the claimed invention is not rendered inoperative by the support element structure of the claimed invention and its integral electrical connection therewith, as asserted by the Examiner. One skilled in the art in reading Applicant's Specification will readily understand that which Applicant is claiming as the invention (in accordance with the amended claims herein) and how to make and/or use the claimed invention.

With respect to the rejection of claim 12 under 35 USC § 112, first paragraph, the Examiner's attention is respectfully directed to at least Applicant's Specification page 4, line 20 through page 5, line 18; and page 6, line 17 through page 7, line 14, which provide detailed support and description on the design (e.g., the dimensional height ("H") and width ("W") characteristics of the transmission line/element, and the spacing ("D") and length ("L") characteristics of the support elements/arms), function and material of the claimed support elements and transmission element thereby enabling one skilled in the art to make and use the claimed invention without resorting to undue experimentation.

In view of the above, Applicant respectfully requests the withdrawal of the outstanding rejections of the aforementioned claims under 35 USC § 112, first paragraph.

<u>Rejection of Claims under 35 USC § 102(b) and 35 USC § 102(e)</u>

The Office Action rejected claims 1, 2 and 5 under 35 USC § 102(b) as being anticipated by U.S. Patent No. 4,340,873 issued to E. Bastida (hereinafter "Bastida"), and

claims 1, 2, 4-8, 10-11, 14, 15, and 17 as being anticipated by U.S. Patent No. 5,594,393 issued to W. Bischof (hereinafter "Bischof"). The Office Action also rejected claims 1, 2 and 5 under 35 USC § 102(e) as being anticipated by U.S. Patent No. 6,714,104 issued to O. Salmela (hereinafter "Salmela"), and claims 1, 2, 5-9 and 14-16 as being anticipated by U.S. Patent No. 6,603,376 issued to M. Handforth et al. (hereinafter "Handforth").

Applicant respectfully disagrees with each of the aforementioned rejections under 35 USC § 102(b) and 35 USC § 102(e) for the reasons set forth below. Further, Applicant has amended the currently independent claims to even make clearer and more particularly claim that which Applicant regards as the invention.

In brief, Applicant does not understand any of the above-cited references (i.e., Bastida, Bischof, Salmela and Handforth) to teach or suggest (1) suspending the claimed transmission element a desired distance away from the ground plane; <u>and</u> (2) electrically connecting such transmission element to the ground plane such that the transmission line is not in electrical contact with a substrate other than through the support elements, as required by Applicant's amended independent claims herein. That is, Applicant does not find a teaching or suggestion in Bastida, Bischof, Salmela or Handforth of a direct electrical connection between the ground layer and the signal trace (i.e., transmission element/line) such that the operation of the transmission line at microwave frequencies essentially eliminates the supporting arms from a signal perspective, as already explained in greater detail hereinabove.

That is, the transmission line and supporting arm structure of the present invention substantially reduces the signal attenuation of high-frequency RF signals propagating along the transmission line. This reduction is the result of separating the transmission line from the substrate and electrically connecting such transmission line to the ground plane such that the transmission line is not in electrical contact with the substrate other than through the support elements. Accordingly, the exposure of the propagating signal to any electromagnetic field present in the substrate is reduced thereby substantially reducing, or eliminating, the signal attenuation.

As will be appreciated by one skilled in the art, the term "electrically connected" refers to the fact that a direct conducting path is provided between two conductors such that free electrons can freely move from one conductor to the other. A potential

difference and, therefore, electric fields cannot exist between directly electrically connected points (see, e.g., the aforementioned "Microwave Engineering" textbook at page 21; and R. Collin, Foundation For Microwave Engineering, McGraw Hill, 1992, pp. 72-74, copy of each enclosed for reference). Such an "electrical connection" plays an important role in Applicant's claimed invention, in particular, an electrical connection that electrically connects the transmission element to the ground plane such that the transmission line is not in electrical contact with a substrate other than through the support elements.

Therefore, as to Bastida, Applicant understands that Bastida to use a nonconductive silicon dioxide strip material between the central conductor and other conducting elements (see, e.g., Bastida, Figures 4 and 5). Therefore, there will not be a direct electron transfer between the two conductors (see, Bastida, Figure 4) as such there is not an "electrical connection". As to Salmela, the signal cable and ground cable are not in physical/electrical contact (see, e.g., Salmela, column 4, line 30). As to Handforth, at column 4, line 37, Handforth states that the stripline and optional ground planes are spaced by an air channel and any known dielectric material. Air (as any other dielectric material) is not conductive and does not constitute the electrical connection required by Applicant's claimed invention. As to Bischof, at column 4, lines 4-8, Bischof states that the posts consist either of a dielectric material or of a conducting material. As already indicted, a dielectric material is not conductive. In the later case, it is necessary for them to be insulated from the metallized conducting track on the substrate, as such, the insulation does not constitute Applicant's claimed electrical connection.

Regarding the rejection of each of the pending dependent claims these claims depend ultimately from one of the pending amended independent claims herein which Applicant submits are patentably distinct over Bastida, Bischof, Salmela and Handforth for the aforesaid reasons. Therefore, as the pending dependent claims herein contain all the limitations of the pending amended independent claims from which they depend, Applicant respectfully submits that these dependent claims are also patentably distinct over Bastida, Bischof, Salmela and Handforth, for the aforesaid reasons, as well as other elements these claims add in combination to their base claim.

In view of the foregoing, Applicant respectfully submits that each of the currently pending claims in the application is in condition for allowance and reconsideration is requested.

Favorable action is respectfully requested. Should the Examiner believe anything further is desirable in order to place the application in even better condition for allowance, the Examiner is invited to contact the undersigned at the telephone number listed below.

Respectfully submitted,

Carsten Metz

By

Donald P. Dinella
Attorney for Applicants
Reg. No. 39,961
908-582-8582

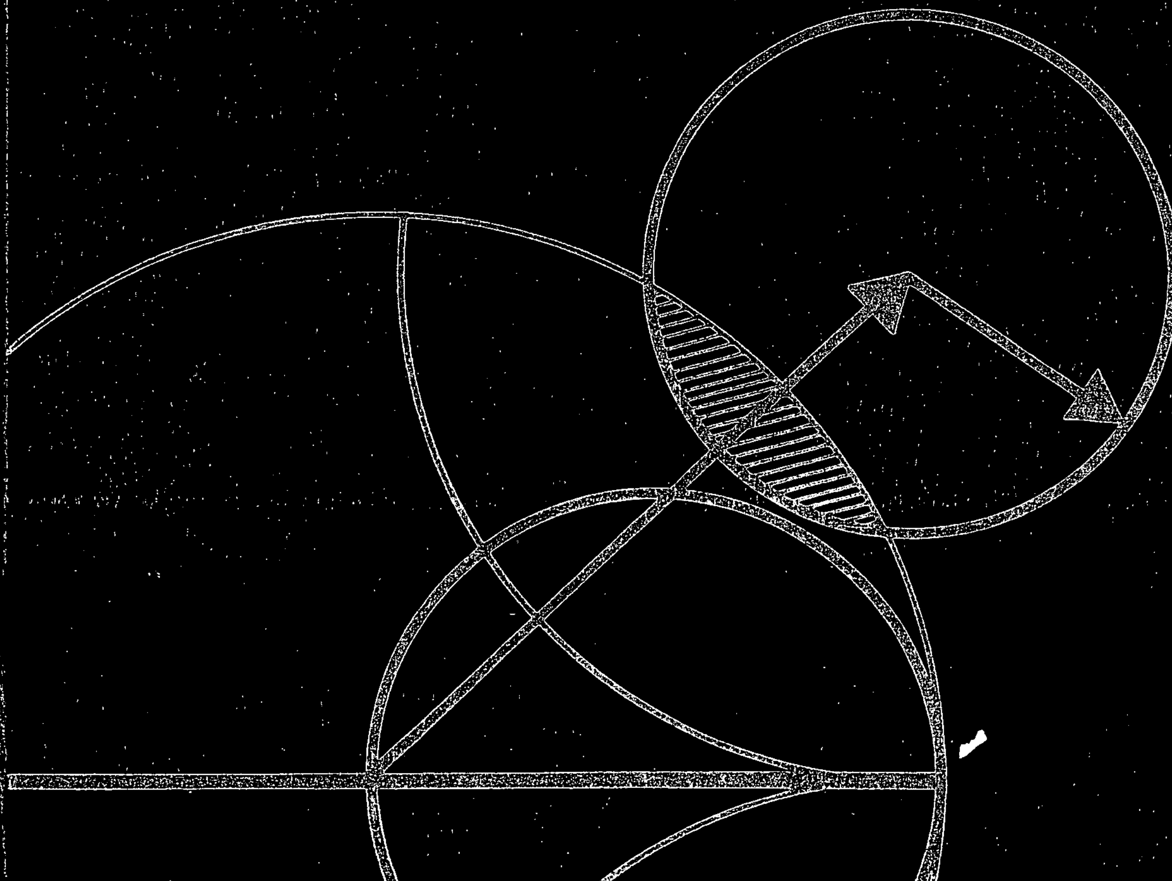
Date: September 15, 2005

Docket Administrator (Room 3J-219)
Lucent Technologies Inc.
101 Crawfords Corner Road
Holmdel, NJ 07733-3030

Attachments:
(4) Sheets of Formal Drawings
D. M. Pozar, Microwave Engineering, Addison Wesley 1990, pp. 1-2, 21, and 79-80
R. Collin, Foundations For Microwave Engineering, McGraw Hill, 1992, pp. 72-74

Microwave Engineering

David M. Pozar



CHAPTER 1

Introduction

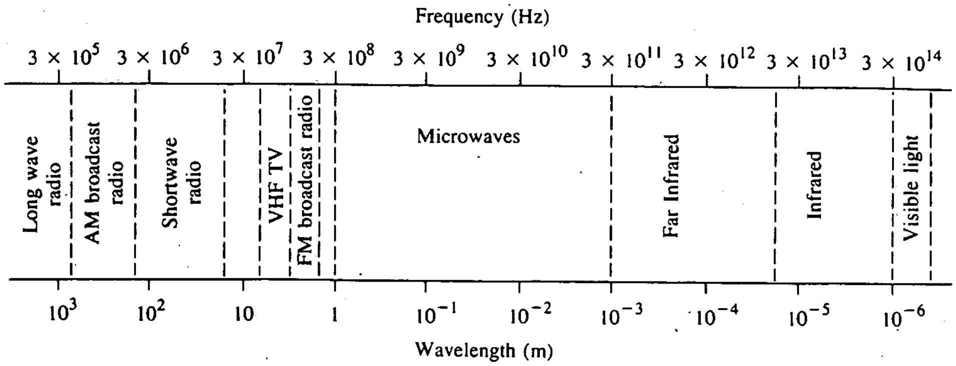
This chapter is intended to provide a brief overview of microwave engineering, including some of the major applications of microwaves. Also included is a short history of the microwave field. The student should thus get a glimpse of the larger context of the material that follows.

1.1 WHAT IS MICROWAVE ENGINEERING?

The term *microwave* refers to alternating current signals with frequencies between 300 MHz (3×10^8 Hz) and 300 GHz (3×10^{11} Hz). See Figure 1.1 for the location of the microwave frequency band in the electromagnetic spectrum. The period, $T = 1/f$, of a microwave signal then ranges from 3 ns (3×10^{-9} sec) to 3 ps (3×10^{-12} sec), respectively, and the corresponding electrical wavelength ranges from $\lambda = c/f = 1$ m to $\lambda = 1$ mm, respectively, where $c = 3 \times 10^8$ m/sec, the speed of light in a vacuum. Signals with wavelengths on the order of millimeters are called *millimeter* waves. It is really the values of the above quantities that make microwave engineering different from other areas of electrical engineering. Because of the high frequencies (and short wavelengths), standard circuit theory cannot be used directly to solve microwave network problems. In a sense, standard circuit theory is an approximation or special case of the broader theory of electromagnetics as described by Maxwell's equations. This is due to the fact that, in general, the lumped circuit element approximations of circuit theory are not valid at microwave frequencies. Microwave components usually are *distributed* elements, where the phase of a voltage or current changes signficantly over the physical length of the device because the device dimensions are on the order of the microwave wavelength. At much lower frequencies, the wavelength is so large that there is little variation in phase across the dimensions of a component.

At the other extreme is optical engineering, where the wavelength is much shorter than the dimensions of the components. In this case Maxwell's equations can be simplified to the geometrical optics regime, and optical systems can be designed with the theory of geometrical optics. Such techniques are sometimes even applicable to millimeter wave systems, where they are referred to as "quasi-optical."

In microwave engineering, then, one must begin with Maxwell's equations and their solutions. It is in the nature of these equations that mathematical complexity arises, as Maxwell's equations involve vector differential or integral operations on vector field



Typical Frequencies

AM broadcast band	535–1605 kHz
Shortwave radio	3–30 MHz
FM broadcast band	88–108 MHz
VHF TV (2–4)	54–72 MHz
VHF TV (5–6)	76–88 MHz
UHF TV (7–13)	174–216 MHz
UHF TV (14–83)	470–890 MHz
Microwave ovens	2.45 GHz

Approximate Band Designations

L-band	1–2 GHz
S-band	2–4 GHz
C-band	4–8 GHz
X-band	8–12 GHz
Ku-band	12–18 GHz
K-band	18–26 GHz
Ka-band	26–40 GHz
U-band	40–60 GHz

FIGURE 1.1 The electromagnetic spectrum.

quantities, and these fields are functions of spatial coordinates. One of the goals of this book, however, is to try to reduce the complexity of a field theory solution to a result which can be expressed in terms of circuit theory. The field theory solution generally provides a complete description of the electromagnetic field at every point in space—usually much more information than we really need. We usually are interested in terminal quantities, such as power, impedance, voltage, current, etc., which can often be expressed in terms of circuit theory concepts. This complexity adds to the difficulty, as well as to the challenge and reward, of microwave engineering.

1.2 APPLICATIONS OF MICROWAVES

Just as the high frequencies and small wavelengths of microwave energy make for difficult analysis and design of microwave components, these same factors provide unique applications for microwave systems. This is because of the following considerations:

- Antenna gain is proportional to the electrical size of the antenna. At higher frequencies, more antenna gain is possible for a given physical antenna size.
- More bandwidth (and hence information-carrying capacity) can be realized at higher frequencies. A 1% bandwidth at 600 MHz is 6 MHz (the bandwidth of one television channel), while at 60 GHz a 1% bandwidth is 600 MHz (about 100 television channels).

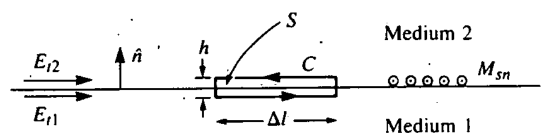


FIGURE 2.5 Closed contour C for equation (2.33).

Fields at a Dielectric Interface

At an interface between two lossless dielectric materials, no charge or surface current densities will ordinarily exist. Equations (2.31), (2.32), (2.36), and (2.37) then reduce to

$$\hat{n} \cdot \bar{D}_1 = \hat{n} \cdot \bar{D}_2, \tag{2.38a}$$

$$\hat{n} \cdot \bar{B}_1 = \hat{n} \cdot \bar{B}_2, \tag{2.38b}$$

$$\hat{n} \times \bar{E}_1 = \hat{n} \times \bar{E}_2, \tag{2.38c}$$

$$\hat{n} \times \bar{H}_1 = \hat{n} \times \bar{H}_2. \tag{2.38d}$$

In words, these equations state that the normal components of $\bar{D}$ and $\bar{B}$ are continuous across the interface, and the tangential components of $\bar{E}$ and $\bar{H}$ are equal across the interface. Because Maxwell's equations are not all linearly independent, the six boundary conditions contained in the above equations are not all linearly independent. Thus, the enforcement of (2.38c) and (2.38d) for the four tangential field components, for example, will automatically force the satisfaction of the equations for the continuity of the normal components.

Fields at the Interface with a Perfect Conductor (Electric Wall)

Many problems in microwave engineering involve boundaries with good conductors (e.g., metals), which can often be assumed as lossless ($\sigma \to \infty$). In this case of a perfect conductor, all field components must be zero inside the conducting region. This result can be seen by considering a conductor with finite conductivity ($\sigma < \infty$) and noting that the skin depth (the depth to which most of the microwave power penetrates) goes to zero as $\sigma \to \infty$. (Such an analysis will be performed in Section 2.7.) If we also assume here that $\bar{M}_s = 0$, which would be the case if the perfect conductor filled all the space on one side of the boundary, then (2.31), (2.32), (2.36), and (2.37) reduce to the following:

$$\hat{n} \cdot \bar{D} = \rho_s, \tag{2.39a}$$

$$\hat{n} \cdot \bar{B} = 0, \tag{2.39b}$$

$$\hat{n} \times \bar{E} = 0, \tag{2.39c}$$

$$\hat{n} \times \bar{H} = \bar{J}_s, \tag{2.39d}$$

where ρ_s and $\bar{J}_s$ are the electric surface charge density and current density, respectively, on the interface, and $\hat{n}$ is the normal unit vector pointing out of the perfect conductor. Such a boundary is also known as an *electric wall*, since the tangential components of $\bar{E}$ are "shorted out," as seen from (2.39c), and must vanish at the surface of the conductor.

The reflection coefficient of (3.35) was defined as the ratio of the reflected to the incident voltage wave amplitudes at the load ($\ell = 0$), but this quantity can be generalized to any point ℓ on the line as follows. From (3.34a), with $z = -\ell$, the ratio of the reflected component to the incident component is

$$\Gamma(\ell) = \frac{V_o^- e^{-j\beta\ell}}{V_o^+ e^{j\beta\ell}} = \Gamma(0)e^{-2j\beta\ell}, \tag{3.42}$$

where $\Gamma(0)$ is the reflection coefficient at $z = 0$, as given by (3.35). This form is useful when transforming the effect of a load mismatch down the line.

We have seen that the real power flow on the line is a constant, but that the voltage amplitude, at least for a mismatched line, is oscillatory with position on the line. The perceptive reader may therefore have concluded that the impedance seen looking into the line must vary with position, and this is indeed the case. At a distance $\ell = -z$ from the load, the input impedance seen looking toward the load is

$$Z_{\text{in}} = \frac{V(-\ell)}{I(-\ell)} = \frac{V_o^+\left[e^{j\beta\ell} + \Gamma e^{-j\beta\ell}\right]}{V_o^+\left[e^{j\beta\ell} - \Gamma e^{-j\beta\ell}\right]} Z_0 = \frac{1 + \Gamma e^{-2j\beta\ell}}{1 - \Gamma e^{-2j\beta\ell}} Z_0, \tag{3.43}$$

where (3.36a,b) have been used for $V(z)$ and $I(z)$. A more usable form may be obtained by using (3.35) for Γ in (3.43):

$$\begin{aligned} Z_{\text{in}} &= Z_0 \frac{(Z_L + Z_0)e^{j\beta\ell} + (Z_L - Z_0)e^{-j\beta\ell}}{(Z_L + Z_0)e^{j\beta\ell} - (Z_L - Z_0)e^{-j\beta\ell}} \\ &= Z_0 \frac{Z_L \cos\beta\ell + jZ_0 \sin\beta\ell}{Z_0 \cos\beta\ell + jZ_L \sin\beta\ell} \\ &= Z_0 \frac{Z_L + jZ_0 \tan\beta\ell}{Z_0 + jZ_L \tan\beta\ell}. \end{aligned} \tag{3.44}$$

This is an important result giving the input impedance of a length of transmission line with an arbitrary load impedance. We will refer to this result as the transmission line impedance equation; some special cases will be considered next.

Special Cases of Lossless Terminated Lines

A number of special cases of lossless terminated transmission lines will frequently appear in our work, so it is appropriate to consider the properties of such cases here.

Consider first the transmission line circuit shown in Figure 3.5, where a line is terminated in a short circuit, $Z_L = 0$. From (3.35) it is seen that the reflection coefficient for a short circuit load is $\Gamma = -1$; it then follows from (3.41) that the standing wave ratio is infinite. From (3.36) the voltage and current on the line are

$$V(z) = V_o^+\left[e^{-j\beta z} - e^{j\beta z}\right] = -2jV_o^+ \sin\beta z, \tag{3.45a}$$

$$I(z) = \frac{V_o^+}{Z_0}\left[e^{-j\beta z} + e^{j\beta z}\right] = \frac{2V_o^+}{Z_0}\cos\beta z, \tag{3.45b}$$

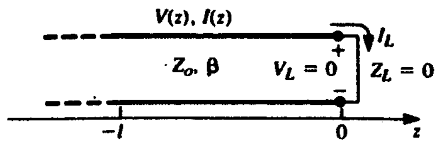


FIGURE 3.5 A transmission line terminated in a short circuit.

which shows that $V = 0$ at the load (as it should, for a short circuit), while the current is a maximum there. From (3.44), or the ratio $V(-\ell)/I(-\ell)$, the input impedance is

$$Z_{\text{in}} = jZ_0 \tan \beta \ell, \tag{3.45c}$$

which is seen to be purely imaginary for any length, ℓ, and to take on all values between $+j\infty$ and $-j\infty$. For example, when $\ell = 0$ we have $Z_{\text{in}} = 0$, but for $\ell = \lambda/4$ we have $Z_{\text{in}} = \infty$ (open-circuit). Equation (3.45c) also shows that the impedance is periodic in ℓ, repeating for multiples of $\lambda/2$. The voltage, current, and input reactance for the short-circuited line are plotted in Figure 3.6.

Next consider the open-circuited line shown in Figure 3.7, where $Z_L = \infty$. Dividing the numerator and denominator of (3.35) by Z_L and allowing $Z_L \to \infty$ shows that the reflection coefficient for this case is $\Gamma = 1$, and the standing wave ratio is again infinite. From (3.36) the voltage and current on the line are

$$V(z) = V_o^+ \left[e^{-j\beta z} + e^{j\beta z}\right] = 2V_o^+ \cos \beta z, \tag{3.46a}$$

$$I(z) = \frac{V_o^+}{Z_0} \left[e^{-j\beta z} - e^{j\beta z}\right] = \frac{-2jV_o^+}{Z_0} \sin \beta z, \tag{3.46b}$$

which shows that now $I = 0$ at the load, as expected for an open circuit, while the voltage is a maximum. The input impedance is

$$Z_{\text{in}} = -jZ_0 \cot \beta \ell, \tag{3.46c}$$

which is also purely imaginary for any length, ℓ. The voltage, current, and input reactance of the open-circuited line are plotted in Figure 3.8.

Now consider terminated transmission lines with some special lengths. If $\ell = \lambda/2$, (3.44) shows that

$$Z_{\text{in}} = Z_L, \tag{3.47}$$

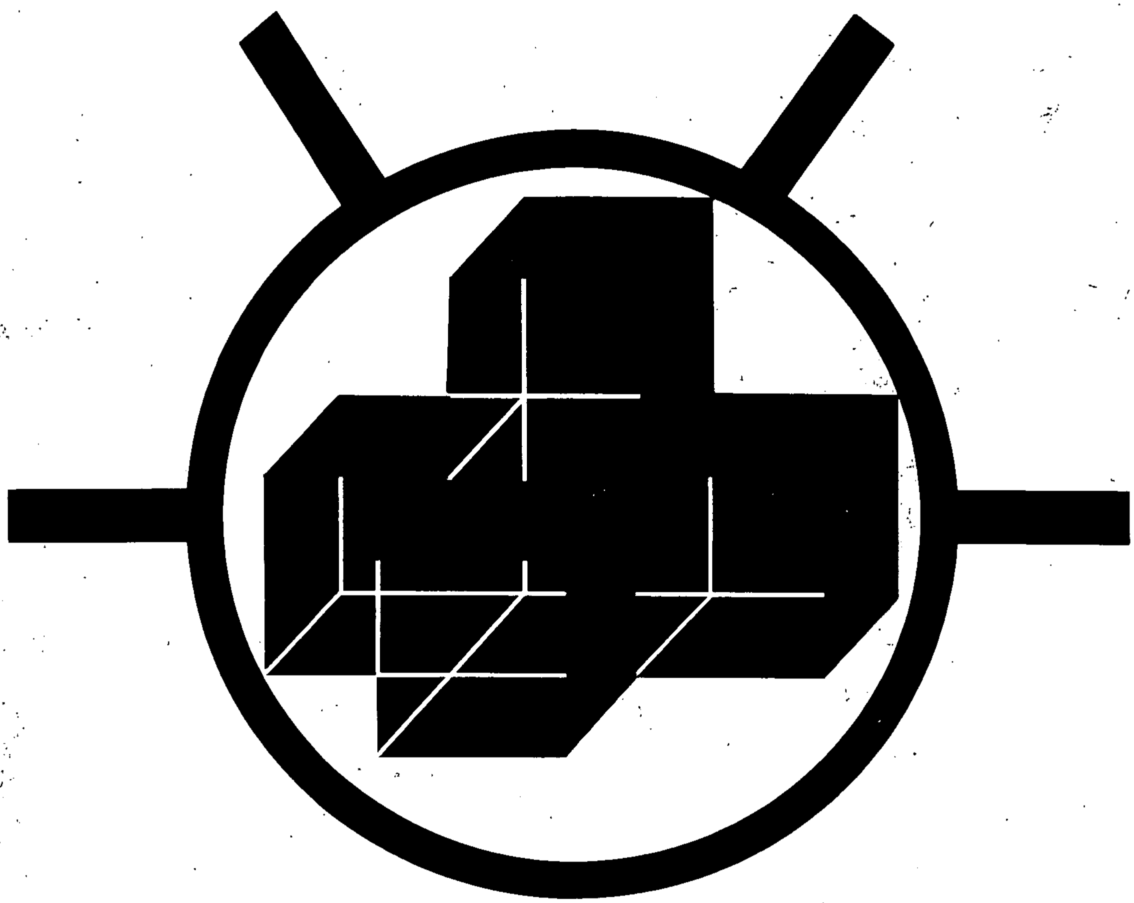
meaning that a half-wavelength line (or any multiple of $\lambda/2$) does not alter or transform the load impedance, regardless of the characteristic impedance.

If the line is a quarter-wavelength long or, more generally, $\ell = \lambda/4 + n\lambda/2$, for $n = 1, 2, 3 \cdots$, (3.44) shows that the input impedance is given by

$$Z_{\text{in}} = \frac{Z_0^2}{Z_L}. \tag{3.48}$$

SECOND EDITION

FOUNDATIONS FOR MICROWAVE ENGINEERING



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Part 3 presents the theory for waves in hollow rectangular and circular waveguides (pipes). In the beginning section of Part 2, we show that Maxwell's equations can be separated into equations that describe three types of waves. These are transverse electromagnetic waves (TEM), transverse electric (TE), and transverse magnetic (TM) waves. The TEM wave is the principal wave that can exist on a transmission line. The TE and TM waves are characterized by having no axial component of electric and magnetic field respectively. The TE and TM waves are the fundamental wave types that can exist in hollow-pipe waveguides. Hollow-pipe waveguides do not support TEM waves. The ability to reduce Maxwell's equations into three set of equations, one set for each wave type, facilitates the analysis of transmission lines and waveguides. Thus this decomposition of Maxwell's equations is carried out in the first section of Part 2.

PART 1
WAVES ON TRANSMISSION LINES

In this section we introduce the topic of voltage and current waves on a two-conductor transmission line by using a distributed-circuit model of the transmission line. This allows us to explore a number of fundamental properties of one-dimensional waves without having to consider the electromagnetic fields in detail. The distributed-circuit-model approach has limitations and in general must be replaced by a detailed solution for the electromagnetic field associated with the guiding structure if we want to determine the distributed-circuit parameters. The field analysis of transmission lines is presented in Part 2.

3.1 WAVES ON AN IDEAL TRANSMISSION LINE

In Fig. 3.1a we show a two-conductor transmission line consisting of two parallel round conductors (wires). The conductors will be assumed to be perfect, i.e., have infinite conductivity. The conductors extend from $z = 0$ to infinity, thus forming a semiinfinite transmission line. At $z = 0$ a voltage generator with internal resistance R_g is connected to the transmission line. The generator produces a voltage $\mathscr{V}_g(t)$ that is impressed across the transmission line. If the generator is switched on at time $t = 0$, a current $\mathscr{I}(t)$ will begin to flow into the upper conductor. A return current $-\mathscr{I}(t)$ must then flow on the lower conductor since current flow through the generator must be continuous. The return current is produced by the action of the electric field established between the two conductors. Since the transmission line is semiinfinite in length, there is no direct conducting path between the upper and lower conductors. However, there is a distributed capacitance C

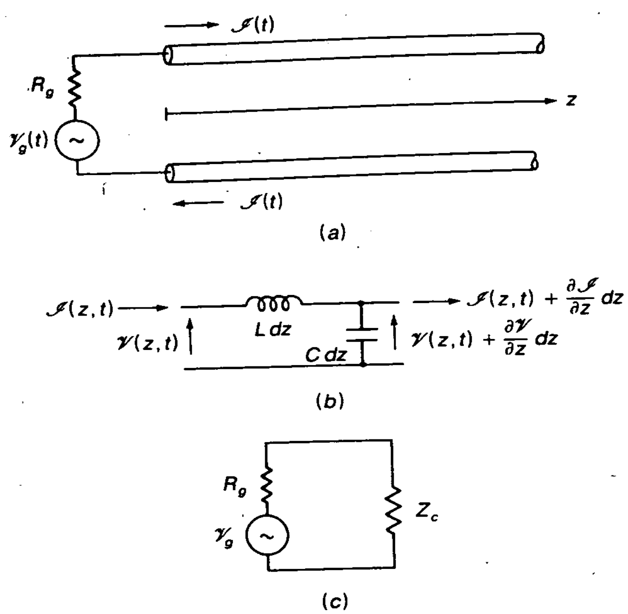

FIGURE 3.1
(*a*) An ideal two-conductor transmission line connected to a voltage generator; (*b*) equivalent circuit of a differential section of the transmission line with no loss; (*c*) equivalent circuit seen by the generator.

per meter between the two conductors; so we have a capacitive or displacement current flowing from the upper conductor to the lower conductor.

The electric current results in a magnetic field around the conductors and consequently the transmission line will also have a distributed series inductance L per meter. We can model a differential section dz of this transmission line as a series inductance $L\,dz$ and a shunt capacitance $C\,dz$ as shown in Fig. 3.1*b*. If the conductors had finite conductivity, we would also need to include a series resistance in the equivalent circuit of a differential section. However, we are assuming that the conductors are perfect; so the series distributed resistance R per meter is zero.

Since electrical effects propagate with a finite velocity v (the speed of light in vacuum), it should be clear that the voltage $\mathscr{V}(z,t)$ and current $\mathscr{I}(z,t)$ at some arbitrary point z on the transmission line will be zero until a time z/v has elapsed after switching the generator on. We will show that the generator launches voltage and current waves on the transmission line that propagate with a finite velocity. The equations that describe these waves are established by applying Kirchhoff's circuit laws to the equivalent circuit of a differential section of the transmission line, along with a specification of the terminal relationships (boundary conditions) that must hold at the generator end.

At some arbitrary point z on the transmission line, let the voltage and current be given by $\mathscr{V}(z,t), \mathscr{I}(z,t)$. At a differential distance dz further

along, the voltage and current have changed by small amounts $(\partial \mathscr{V}/\partial z)\,dz$ and $(\partial \mathscr{I}/\partial z)\,dz$; so the output voltage and current at $z + dz$ will be

$$\mathscr{V}(z+dz,t) = \mathscr{V}(z,t) + \frac{\partial \mathscr{V}(z,t)}{\partial z}\,dz$$

$$\mathscr{I}(z+dz,t) = \mathscr{I}(z,t) + \frac{\partial \mathscr{I}(z,t)}{\partial z}\,dz$$

The sum of all potential drops around the circuit must be zero; so we have

$$-\mathscr{V} + L\,dz\frac{\partial \mathscr{I}}{\partial t} + \mathscr{V} + \frac{\partial \mathscr{V}}{\partial z}\,dz = 0$$

or

$$\frac{\partial \mathscr{V}(z,t)}{\partial z} = -L\frac{\partial \mathscr{I}(z,t)}{\partial t} \tag{3.1a}$$

The sum of currents flowing into the output node must also be zero; so we can write

$$\mathscr{I} - C\,dz\frac{\partial \mathscr{V}}{\partial t} - \mathscr{I} - \frac{\partial \mathscr{I}}{\partial z}\,dz = 0$$

or

$$\frac{\partial \mathscr{I}(z,t)}{\partial z} = -C\frac{\partial \mathscr{V}(z,t)}{\partial t} \tag{3.1b}$$

These two partial differential equations describe the relationship between the voltage and current waves on the transmission line.

We can obtain an equation for the voltage $\mathscr{V}(z,t)$ by differentiating (3.1a) with respect to z and using (3.1b) to eliminate the current; thus

$$\frac{\partial^2 \mathscr{V}(z,t)}{\partial z^2} = -L\frac{\partial^2 \mathscr{I}}{\partial z\,\partial t} = -L\left(-C\frac{\partial^2 \mathscr{V}}{\partial t^2}\right)$$

or

$$\frac{\partial^2 \mathscr{V}(z,t)}{\partial z^2} - LC\frac{\partial^2 \mathscr{V}(z,t)}{\partial t^2} = 0 \tag{3.2a}$$

In a similar way we obtain

$$\frac{\partial^2 \mathscr{I}(z,t)}{\partial z^2} - LC\frac{\partial^2 \mathscr{I}(z,t)}{\partial t^2} = 0 \tag{3.2b}$$

The product LC has the dimensions of one over velocity squared. These two equations are one-dimensional wave equations and describe waves propagating with a velocity†

$$v = \frac{1}{\sqrt{LC}} \tag{3.3}$$

†For an ideal transmission line in air, $v = c = 3 \times 10^8$ m/s, the velocity of light.

AMENDMENTS TO THE DRAWINGS

In response to the objections to the drawings as set forth in the above-identified Office Action, Applicant has submitted four (4) sheets of corrected formal drawings and requests that such corrected formal drawings be made of record in the subject application. Applicant respectfully submits that the corrected drawings herein overcome the objections to the drawings set forth in the Office Action, which objections Applicant requests be withdrawn.

Each of the drawings sheets submitted herein is identified in the top margin as a "Replacement Sheet", as indicated in the Office Action.

Attachment: Replacement Sheets.